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(54) **Process and apparatus for performing initial carrier frequency offset in an OFDM communication system**

(57) A process for estimating the carrier frequency offset (CFO) for a Orthogonal Frequency Division Multiplex (OFDM) communication system, said process being performed in a receiver receiving a pilot sequence $\mathbf{x} = (x_0, x_1, x_2, \dots, x_{N-1})^T$, said process involving the step of determining a value of θ which maximizes the formula:

$$C(\theta) = -\mathbf{d}_\theta^H \mathbf{Y}^H \mathbf{X} (\mathbf{X}^H \mathbf{X} + \sigma^2 \mathbf{Q}^{-1})^{-1} \frac{1}{\sigma^2} \mathbf{X}^H \mathbf{Y} \mathbf{d}_\theta$$

where

$$\mathbf{X} = \begin{pmatrix} x_0 & x_{N-1} & \cdots & x_{N-L-1} \\ x_1 & x_0 & \cdots & x_{N-L-2} \\ \vdots & \vdots & \ddots & \vdots \\ x_{L-2} & x_{L-3} & \cdots & x_{N-1} \\ x_{L-1} & x_{L-2} & \cdots & x_0 \\ \vdots & \vdots & \ddots & \vdots \\ x_{N-1} & x_{N-2} & \cdots & x_{N-L} \end{pmatrix}$$

$$\mathbf{d}_\theta = (1, e^{2\pi \frac{i\theta}{N}}, \dots, e^{2\pi \frac{(N-1)i\theta}{N}})^T;$$

$\mathbf{Y}$ is the diagonal matrix of main diagonal $\mathbf{y}$;

σ^2 is the noise power and $\mathbf{Q}$ is the channel time covariance matrix equal to $1/L \mathbf{I}_L$, with $\mathbf{I}_L$ being the $L \times L$ identity matrix and L corresponding to the presumed length of the channel.

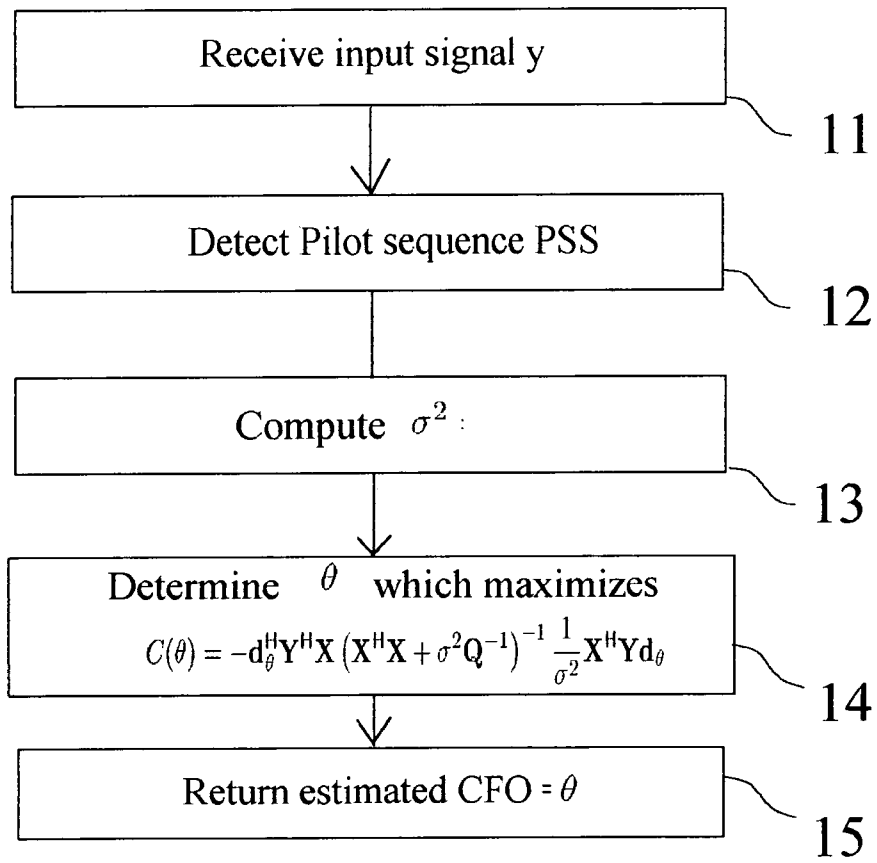


Fig. 1

Description**Technical field**

5 **[0001]** The invention relates to wireless digital communication and particularly to a process for performing the estimation of a Carrier Frequency Offset in a OFDM communication system, during the initial synchronization steps (i.e. prior to channel estimation).

Background Art

10 **[0002]** The Orthogonal Frequency Division Multiplexing (OFDM) modulation is a widely developing technique in the field of wireless digital communications thanks to the high possibilities offered by simplified digital signal processing based on Discrete Fourier Transform computations.

15 **[0003]** However, before a signal being received by a User Equipment (UE) may be properly processed, a preliminary but critical synchronization step has to be achieved including the estimation of the so-called Carrier Frequency Offset (CFO). Indeed, because of the mismatches of the internal oscillators and also because of the Doppler effect resulting from mobile communications, the frequencies at the base station (eNodeB) and at the receiving terminal (UE) may differ, and such a difference might be detrimental to the targeted transmission rate. In particular, if the CFO problem is not properly corrected, it is hazardous to perform the Discrete Fourier Transform stage at the receiver. In many systems, 20 such as 3GPP-LTE, a first rough synchronization step is performed thanks to a primary synchronization sequence (PSS) which at this point is the only known pilot sequence accessible to the receiver.

[0004] Before performing the next synchronization steps which require to use the DFT block at the reception, it is critical to have a good CFO estimate; otherwise the performance of the subsequent synchronization steps might be heavily impaired.

25 **[0005]** A first, very well known, technique for estimating the CFO is disclosed by P. H. Moose in "A technique for orthogonal frequency-division multiplexing frequency offset correction", in IEEE Trans. on Communications, vol. 42, no. 10, pp. 2908-2914, Oct. 1994. The technique is based on the use of a specific pilot sequence, a pilot OFDM symbol composed of two identical vectors.

30 **[0006]** Such a technique presents the significant advantage of being workable without the knowledge of the channel characteristics and, therefore, is highly useful during the first or coarse synchronization process when only limited information is available to the receiver. The clear drawback of this well-known technique lies in the fact that the repetitive transmission of the Moose sequence is a large waste in terms of spectral efficiency since this sequence is not in general reusable to other synchronization purposes.

35 **[0007]** For such a reason in particular, the use of the Moose sequence has been excluded in some standards, such as the 3GPP-LTE for instance, and there is only provided the so-called Primary Synchronization Sequence (PSS) for achieving the estimation of the CFO. In theory, by means of a systematic scanning of a dense discrete set of possible central frequencies, it is feasible to evaluate the CFO from the detection of the PSS but such a systematic scanning would require a great amount of digital processing resources and thus increase the processing costs of the receiver.

40 **[0008]** It is therefore desirable to have an alternative technique which can achieve CFO estimation with limited digital processing resources.

Summary of the invention

45 **[0009]** It is an object of the present invention to provide a process for estimating the Carrier Frequency Offset (CFO) between a receiver and a transmitter in an OFDM communication system.

[0010] It is another object of the present invention to provide a process and apparatus which achieves CFO estimation on the basis of any predetermined pilot sequence.

[0011] It is still another object of the present invention to provide a process and apparatus for performing CFO estimation on the basis of the so-called Primary Synchronization Signal (PSS).

50 **[0012]** It is still another object of the present invention to provide a process and apparatus for performing CFO estimation for a Long Term Evolution (LTE) communication system.

[0013] Those and other objects are achieved by means of the process according to the present invention which is performed in a communication device receiving a known (pilot) sequence $\mathbf{x} = (x_0, x_1, x_2, \dots, x_{N-1})^T$ of length N.

[0014] The process involves the determination of a value of θ which maximizes the formula:

$$C(\theta) = -\mathbf{d}_\theta^H \mathbf{Y}^H \mathbf{X} (\mathbf{X}^H \mathbf{X} + \sigma^2 \mathbf{Q}^{-1})^{-1} \frac{1}{\sigma^2} \mathbf{X}^H \mathbf{Y} \mathbf{d}_\theta$$

where

$$\mathbf{X} = \begin{pmatrix} x_0 & x_{N-1} & \cdots & x_{N-L-1} \\ x_1 & x_0 & \cdots & x_{N-L-2} \\ \vdots & \vdots & \ddots & \vdots \\ x_{L-2} & x_{L-3} & \cdots & x_{N-1} \\ x_{L-1} & x_{L-2} & \cdots & x_0 \\ \vdots & \vdots & \ddots & \vdots \\ x_{N-1} & x_{N-2} & \cdots & x_{N-L} \end{pmatrix}$$

$$\mathbf{d}_\theta = (1, e^{2\pi \frac{i\theta}{N}}, \dots, e^{2\pi \frac{(N-1)i\theta}{N}})^T;$$

$\mathbf{Y}$ is the diagonal matrix of main diagonal $\mathbf{y}$;

σ^2 is the noise power and $\mathbf{Q}$ is the assumed channel time covariance matrix.

[0015] In one embodiment, the $\mathbf{Q}$ matrix is set to be equal to $1/L \mathbf{I}_L$, with $\mathbf{I}_L$ the $L \times L$ identity matrix and L corresponding to the presumed length of the channel (i.e. the approximate number of time samples over which the channel spreads).

[0016] Preferably, a dichotomic process is used for the purpose of estimating the value of θ which maximizes $C(\theta)$, such as a steepest descent algorithm

[0017] In one embodiment, the process involves the steps of:

- receiving an input signal $\mathbf{y} = (y_0, \dots, y_{N-1})^T$;
- detecting said pilot signal;
- computing an estimation of the variance σ^2 of the noise;
- setting an estimate of the channel time covariance matrix $\mathbf{Q}$, for instance equal to $1/L \mathbf{I}_L$,
- computing a matrix $\mathbf{A}$ as follows:

$$\mathbf{A} = \mathbf{Y}^H \mathbf{X} (\mathbf{X}^H \mathbf{X} + \sigma^2 \mathbf{Q}^{-1})^{-1} \frac{1}{\sigma^2} \mathbf{X}^H \mathbf{Y}$$

where $\mathbf{X}$ is a pseudo - circulant matrix defined from said pilot signal

$\mathbf{x} = (x_0, x_1, x_2, \dots, x_{N-1})^T$ as follows :

$$\mathbf{X} = \begin{pmatrix} x_0 & x_{N-1} & \cdots & x_{N-L-1} \\ x_1 & x_0 & \cdots & x_{N-L-2} \\ \vdots & \vdots & \vdots & \vdots \\ x_{L-2} & x_{L-3} & \cdots & x_{N-1} \\ x_{L-1} & x_{L-2} & \cdots & x_0 \\ \vdots & \vdots & \vdots & \vdots \\ x_{N-1} & x_{N-2} & \cdots & x_{N-L} \end{pmatrix}$$

$\mathbf{Y}$ is the diagonal matrix of main diagonal $\mathbf{y}$;

σ^2 is the noise power .

- computing the N-1 values of $\hat{a}$ in accordance with the formula:

$$\tilde{a}_k = \sum_{m=1}^{N-k} A_{k+m,m}$$

with A_{ij} the entries of matrix $\mathbf{A}$.

- computing the following vectors:

$$\tilde{\mathbf{b}}_{\Re}^T = \Re[\tilde{a}_1, 2\tilde{a}_2, \dots, (N-1)\tilde{a}_{N-1}]$$

$$\tilde{\mathbf{b}}_{\Im}^T = \Im[\tilde{a}_1, 2\tilde{a}_2, \dots, (N-1)\tilde{a}_{N-1}]$$

- initializing a loop with the two variables:

$$\theta_{\min} = -1/2$$

and

$$\theta_{\max} = 1/2$$

- computing the value of

$$d = \tilde{\mathbf{b}}_{\Re}^T \cdot \mathbf{s}_{\theta} + \tilde{\mathbf{b}}_{\Im}^T \cdot \mathbf{c}_{\theta}$$

with

$$\mathbf{c}_\theta^T = [\cos(2\pi\theta/N), \dots, \cos(2\pi(N-1)\theta/N)]$$

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$$\mathbf{s}_\theta^T = [\sin(2\pi\theta/N), \dots, \sin(2\pi(N-1)\theta/N)]$$

- testing the sign of d ;
- 10 - if d is positive, then updating the value of $\theta_{\min}$ as follows:

$$\theta_{\min} = (\theta_{\min} + \theta_{\max})/2$$

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- if d is negative, then updating the value of $\theta_{\max}$ as follows:

$$\theta_{\max} = (\theta_{\min} + \theta_{\max})/2$$

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- repeating the preceding steps until the completion of the loop and, when the loop completes,
- returning an estimation of said CFO in accordance with the formula:

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$$\hat{\theta} = [\theta_{\min} + \theta_{\max}]/2$$

30 **[0018]** The invention also provides a communication apparatus comprising a receiver for estimating the CFO for an OFDM communication system, which comprises:

- a receiver for receiving an input signal comprising a pilot sequence $\mathbf{x} = (x_0, x_1, x_2, \dots, x_{N-1})^T$,
- means for determining a value of θ which maximizes the formula:

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$$C(\theta) = -\mathbf{d}_\theta^H \mathbf{Y}^H \mathbf{X} (\mathbf{X}^H \mathbf{X} + \sigma^2 \mathbf{Q}^{-1})^{-1} \frac{1}{\sigma^2} \mathbf{X}^H \mathbf{Y} \mathbf{d}_\theta$$

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where

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$$\mathbf{X} = \begin{pmatrix} x_0 & x_{N-1} & \cdots & x_{N-L-1} \\ x_1 & x_0 & \cdots & x_{N-L-2} \\ \vdots & \vdots & \vdots & \vdots \\ x_{L-2} & x_{L-3} & \cdots & x_{N-1} \\ x_{L-1} & x_{L-2} & \cdots & x_0 \\ \vdots & \vdots & \vdots & \vdots \\ x_{N-1} & x_{N-2} & \cdots & x_{N-L} \end{pmatrix}$$

$$\mathbf{d}_\theta = (1, e^{2\pi \frac{j\theta}{N}}, \dots, e^{2\pi \frac{(N-1)j\theta}{N}})^T;$$

$\mathbf{Y}$ is the diagonal matrix of main diagonal $\mathbf{y}$;

σ^2 is the noise power and $\mathbf{Q}$ is the channel time covariance matrix.

[0019] The invention is particularly adapted for the estimation of the CFO for a 3GPP-LTE communication network and any type of mobile OFDM-based network with no CFO-dedicated synchronization sequence.

Description of the drawings

[0020] Other features of one or more embodiments of the invention will best be understood by reference to the following detailed description when read in conjunction with the accompanying drawings.

Figure 1 illustrates a first embodiment of a process in accordance with the present invention.

Figure 2 illustrates a second embodiment of a process in accordance with the present invention.

Figure 3 shows a comparison of the CFO estimates resulting from the traditional Moose technique and the proposed invention, with $N=128$, $L=3$ and $L_{\text{assumed}} \in [3, 9]$

Figures 4-6 show comparative results with the use of the steepest descent algorithm of the invention, and with $N=128$, $L=3$, respectively with 3 to 50 iterations.

Figure 7 illustrates the impact of the choice of the particular pilot sequence for executing the process of the invention

Description of the preferred embodiments

[0021] It will now be described how to estimate the CFO from the sole knowledge of any predetermined pilot sequence, for instance the so-called Primary Synchronization Signal (PSS). However, it should be clear that the PSS is only indicated as an example and that other pilot sequences may be considered by the skilled man.

[0022] In the following, boldface lower-case symbols represent vectors, capital boldface characters denote matrices ($\mathbf{I}_N$ is the $N \times N$ identity matrix). The Hermitian transpose is denoted $(.)^H$. The set of $N \times M$ matrices over the algebra $\mathcal{A}$

is denoted $\mathcal{M}(\mathcal{A}, N, M)$. The operators $\det(\mathbf{X})$ and $\text{tr}(\mathbf{X})$ represent the determinant and the trace of matrix $\mathbf{X}$, respectively. The symbol $E[.]$ denotes expectation.

[0023] Consider a pair of transmitter and receiver communicating through a noisy channel. The transmitter sends a data sequence $\mathbf{x}$ which the receiver captures as a sequence $\mathbf{y}$. The transmission vector channel is denoted $\mathbf{h}$. The noise is modeled as an additive white Gaussian (AWGN) sequence $\mathbf{w}$. The extent of knowledge of the receiver, prior to data transmission, is denoted $\mathbf{I}$. In particular, the receiver frequency reference is not perfectly aligned to that of the transmitter: this introduces a frequency offset θ whose knowledge to the receiver is summarized into the density function $p(\theta|\mathbf{I})$. By inductive reasoning, we provide in the following an expression of the optimal inference the receiver can make on $(\theta|\mathbf{y}, \mathbf{I})$ which we apply to the example of data-aided CFO estimation in OFDM.

[0024] Consider an OFDM system of N subcarriers. The transmitter sends a time-domain pilot sequence $\mathbf{x} = (x_0, \dots, x_{N-1})^T$ (cyclic prefix excluded), received as a sequence $\mathbf{y} = (y_0, \dots, y_{N-1})^T$ (cyclic prefix discarded). The transmission channel is discretized in L taps $\mathbf{h} = (h_0, \dots, h_{L-1})^T$ and the AWGN noise $\mathbf{w} = (w_0, \dots, w_{N-1})^T$ has entries of variance $E[|W_k|^2] = \sigma^2$. For the sake of simplicity, it will not be considered below the information contained in the cyclic prefixes. Let θ represent the CFO to be estimated at the receiver normalized to the subcarrier spacing, i.e. $\theta=1$ is a frequency mismatch of one subcarrier spacing.

[0025] A CFO produces in OFDM a simple phase rotation of all transmitted time-domain symbols x_k of an angle $2\pi k\theta/N$. While it seems feasible to track the CFO in the time domain when the transmitted pilot sequence $\mathbf{x}$ - for instance the PSS - is assumed to be known, it should be noticed that channel estimation is not accessible to the UE during the initial synchronization step, thus preventing direct deciphering of the impact of the channel on the time-domain symbols.

[0026] It is proposed to consider the maximum a posteriori value for θ given the received signal $\mathbf{y}$ defined as

$$\mathbf{y} = \mathbf{H} \mathbf{x} + \mathbf{n} \quad (1)$$

where $\mathbf{H}$ is the circulant matrix of the time-domain OFDM channel (its first row is $\mathbf{h}$) and $\mathbf{n}$ the white Gaussian noise process.
[0027] This classical model can be rewritten

$$\mathbf{y} = \mathbf{X} \mathbf{h} + \mathbf{n} \quad (2)$$

where $\mathbf{h}$ is composed of the L time-domain taps of the channel response and $\mathbf{X}$ is the pseudo-circulant matrix defined as

$$\mathbf{X} = \begin{pmatrix} x_0 & x_{N-1} & \cdots & x_{N-L-1} \\ x_1 & x_0 & \cdots & x_{N-L-2} \\ \vdots & \vdots & \vdots & \vdots \\ x_{L-2} & x_{L-3} & \cdots & x_{N-1} \\ x_{L-1} & x_{L-2} & \cdots & x_0 \\ \vdots & \vdots & \vdots & \vdots \\ x_{N-1} & x_{N-2} & \cdots & x_{N-L} \end{pmatrix}$$

[0028] It is assumed that the CFO is known to be comprised in the set $\theta \in [-1/2, 1/2]$, where θ is normalized to the subcarrier spacing. We want to maximize the probability $p(\theta | y)$.

One may assume uniform prior distribution of $p(\theta)$ in the set $\theta \in [-1/2, 1/2]$, then the maximization problem is concave in the variable θ and therefore can be solved by steepest descent algorithms. After computation, it has been observed that maximizing $p(\theta | y)$ is equivalent to maximize the function $C(\theta)$ defined as

$$C(\theta) = -\mathbf{d}_\theta^H \mathbf{Y}^H \mathbf{X} (\mathbf{X}^H \mathbf{X} + \sigma^2 \mathbf{Q}^{-1})^{-1} \frac{1}{\sigma^2} \mathbf{X}^H \mathbf{Y} \mathbf{d}_\theta$$

where

$$\mathbf{X} = \begin{pmatrix} x_0 & x_{N-1} & \cdots & x_{N-L-1} \\ x_1 & x_0 & \cdots & x_{N-L-2} \\ \vdots & \vdots & \ddots & \vdots \\ x_{L-2} & x_{L-3} & \cdots & x_{N-1} \\ x_{L-1} & x_{L-2} & \cdots & x_0 \\ \vdots & \vdots & \ddots & \vdots \\ x_{N-1} & x_{N-2} & \cdots & x_{N-L} \end{pmatrix}$$

$$\mathbf{d}_\theta = (1, e^{2\pi \frac{i\theta}{N}}, \dots, e^{2\pi \frac{(N-1)i\theta}{N}})^T;$$

$\mathbf{Y}$ is the diagonal matrix of main diagonal $\mathbf{y}$;

σ^2 is the noise power

where $\mathbf{X}$ is the pseudo-circulant matrix defined above, with a first column comprising the any pilot synchronization sequence $\mathbf{x} = (x_0, x_1, x_2, \dots, x_{N-1})^T$ (for instance the PSS), and the next column comprising the circular permutation of the elements of vector $\mathbf{x}$, ie vector $(x_{N-1}, x_0, x_1, \dots, x_{N-2})^T$, and the next one comprising the next consecutive circular permutation $(x_{N-2}, x_{N-1}, x_0, \dots, x_{N-3})^T$ and so on...

[0029] The matrix $\mathbf{Q}$ is the channel time covariance matrix which is assumed to be known. In one particular embodiment, one sets $\mathbf{Q} = 1/L \mathbf{I}_L$, with $\mathbf{I}_L$ being the $L \times L$ identity matrix and L corresponding to the presumed length of the channel. It should be noticed that, generally speaking, L is not known a priori, but it has been advantageously observed that, to some extent, any non-trivial predetermined choice for L (and quite possibly wrong) does not alter much the results and the efficiency of the CFO estimation process. Therefore, the optimal maximum a priori solution simply consists in finding the value θ that maximizes $C(\theta)$.

I. Description of embodiments

[0030] With respect to Figure 1, there is now described the basic steps which are involved in the CFO estimation process in accordance with the present invention.

[0031] The process is executed in any receiver of a OFDM communication system, receiving an input signal $\mathbf{y} = (y_0, \dots, y_{N-1})^T$ in a step 11.

[0032] Then, the process proceeds with a step 12 consisting in the detection of the PSS pilot signal.

[0033] In a step 13, the process computes an estimation of the Signal to Noise Ratio (SNR) and therefore an evaluation of variance of the noise σ^2 . Such evaluation is achieved by techniques and algorithms which are well known to a skilled man and which will not be developed with more details. For instance, the pilot sequence may be used for performing such evaluation.

[0034] In a step 14, the process proceeds with the computation of

$$C(\theta) = -\mathbf{d}_\theta^H \mathbf{Y}^H \mathbf{X} (\mathbf{X}^H \mathbf{X} + \sigma^2 \mathbf{Q}^{-1})^{-1} \frac{1}{\sigma^2} \mathbf{X}^H \mathbf{Y} \mathbf{d}_\theta$$

where

$$\mathbf{X} = \begin{pmatrix} x_0 & x_{N-1} & \cdots & x_{N-L-1} \\ x_1 & x_0 & \cdots & x_{N-L-2} \\ \vdots & \vdots & \ddots & \vdots \\ x_{L-2} & x_{L-3} & \cdots & x_{N-1} \\ x_{L-1} & x_{L-2} & \cdots & x_0 \\ \vdots & \vdots & \ddots & \vdots \\ x_{N-1} & x_{N-2} & \cdots & x_{N-L} \end{pmatrix}$$

$$\mathbf{d}_\theta = (1, e^{2\pi \frac{i\theta}{N}}, \dots, e^{2\pi \frac{(N-1)i\theta}{N}})^T;$$

$\mathbf{Y}$ is the diagonal matrix of main diagonal $\mathbf{y}$;

σ^2 is the noise power and $\mathbf{Q} = 1/L \mathbf{I}_L$.

and identifies the particular value of θ that maximizes $C(\theta)$.

[0035] In one particular embodiment, a processing loop is initiated for the purpose of testing different values of θ and thus identifying the particular value which maximizes $C(\theta)$.

[0036] Alternatively, it has been observed that $C(\theta)$ is concave and therefore a dichotomy algorithm can be advantageously used for achieving a fast computation of the CFO estimation.

[0037] Once determined, the process returns in a step 15 the particular value identified in step 14 as being the estimated CFO .

[0038] As it will be apparent to the skilled man, the process which was described above can be embodied by means of different and numerous algorithms. In addition, it will be clear to the skilled man that the formula above may take various formal presentations showing equivalent computations.

[0039] With respect to figure 2, there will now be described a second embodiment of the invention which requires limited digital processing resources.

[0040] The second embodiment includes steps 21-23 which are identical to steps 11-13 of figure 1.

[0041] Therefore, after the computation of the value of σ^2 , the process proceeds with a step 24 where the value of channel time covariance matrix $\mathbf{Q}$ is being set.

[0042] In one embodiment, the $\mathbf{Q}$ matrix is predetermined. Clearly, the same assumption made in figure 1 may be applicable, for instance $\mathbf{Q} = 1/L \mathbf{I}_L$.

[0043] With the assumption made on matrix $\mathbf{Q}$, the process then proceeds to a step 25 where the following matrix $\mathbf{A}$ (comprising elements $a_{n,m}$) is computed:

$$\mathbf{A} = \mathbf{Y}^H \mathbf{X} (\mathbf{X}^H \mathbf{X} + \sigma^2 \mathbf{Q}^{-1})^{-1} \frac{1}{\sigma^2} \mathbf{X}^H \mathbf{Y}$$

in which

$$\mathbf{X} = \begin{pmatrix} x_0 & x_{N-1} & \cdots & x_{N-L-1} \\ x_1 & x_0 & \cdots & x_{N-L-2} \\ \vdots & \vdots & \vdots & \vdots \\ x_{L-2} & x_{L-3} & \cdots & x_{N-1} \\ x_{L-1} & x_{L-2} & \cdots & x_0 \\ \vdots & \vdots & \vdots & \vdots \\ x_{N-1} & x_{N-2} & \cdots & x_{N-L} \end{pmatrix}$$

where $\mathbf{X}$ is the pseudo-circulant matrix defined above,
 $\mathbf{Y}$ is the diagonal matrix of main diagonal $\mathbf{y}$;
 σ^2 is the noise power.

[0044] In the case of PSS for LTE, the size of the $\mathbf{A}$ matrix is 64x64.

[0045] In one particular embodiment, the process only computes half of matrix $\mathbf{A}$ since only the upper right coefficients $A_{n,m}$, with $n > m$, need to be known for the remaining part of matrix $\mathbf{A}$ as it will be apparent below.

[0046] In a step 26, the process then proceeds with the computation of the $N-1$ values of $\tilde{a}_k$ given by the following formula:

$$\tilde{a}_k = \sum_{m=1}^{N-k} A_{k+m,m}$$

[0047] Then, in a step 27, the process proceeds with the computation of the two following vectors:

$$\begin{aligned} \tilde{\mathbf{b}}_{\Re}^T &= \Re[\tilde{a}_1, 2\tilde{a}_2, \dots, (N-1)\tilde{a}_{N-1}] \\ \tilde{\mathbf{b}}_{\Im}^T &= \Im[\tilde{a}_1, 2\tilde{a}_2, \dots, (N-1)\tilde{a}_{N-1}] \end{aligned}$$

[0048] Then, in a step 28, the process enters into a loop and, in a step 29, initializes the following two variables :

$$\theta_{\min} = -1/2$$

and

$$\theta_{\max} = 1/2$$

[0049] The process then proceeds with a step 30 where the value of d (the derivative of $C(\theta)$ in point θ) is computed:

$$d = \tilde{\mathbf{b}}_{\mathcal{R}}^T \cdot \mathbf{s}_\theta + \tilde{\mathbf{b}}_{\mathcal{I}}^T \cdot \mathbf{c}_\theta$$

with

$$\mathbf{c}_\theta^T = [\cos(2\pi\theta/N), \dots, \cos(2\pi(N-1)\theta/N)]$$

$$\mathbf{s}_\theta^T = [\sin(2\pi\theta/N), \dots, \sin(2\pi(N-1)\theta/N)]$$

[0050] Then, in a step 31, a simple test is performed in order to determine whether d is positive or negative. Indeed, it has been observed that function $C(\theta)$ is concave between $(-\frac{1}{2}, \frac{1}{2})$, what opens the opportunity of a simple test on the sign of d for determining the maximum value of $C(\theta)$.

[0051] If d is found to be positive, then the process proceeds with a step 32 where the value of $\theta_{\min}$ is updated as follows:

$$\theta_{\min} = (\theta_{\min} + \theta_{\max})/2$$

and the process then proceeds to step 34

[0052] Conversely, if d is negative, then the process proceeds to a step 33 where the value of θ is updated as follows:

$$\theta_{\max} = (\theta_{\min} + \theta_{\max})/2$$

[0053] The process then proceeds to a step 34 which is a new test on the end of the loop. If the loop is not terminated, then the process proceeds again to step 28.

[0054] If the loop is terminated, then the process proceeds with a step 35 where the estimated value of the CFO is computed as follows:

$$\hat{\theta} = [\theta_{\min} + \theta_{\max}]/2$$

[0055] The process then completes.

II. SIMULATION AND RESULTS

[0056] In the following, one may consider following we consider an OFDM transmission with $N = 128$ subcarriers. We assume perfect timing offset alignment between the base station and the receiving terminal. A CFO mismatch θ is introduced. The receiver only knows that $\theta \in [-1/2, 1/2]$.

[0057] Figure 3 shows a comparison of the CFO estimates resulting from the traditional Moose technique and the proposed invention, with $N = 128$, $L = 3$ and $L_{\text{assumed}} \in [3, 9]$. One considers a double-half sequence suggested by Moose and the proposed method is compared against the Moose's correlation algorithm on 20,000 channels and CFO realizations (θ is uniformly distributed in $[-1/2, 1/2]$). The channel length is set to $L = 3$, while the a priori on the channel length is either considered known, i.e. $L_{\text{assumed}} = 3$, or wrongly estimated, here $L_{\text{assumed}} = 9$. The respective performances are analyzed in terms of average quadratic error $E[(\hat{\theta} - \theta)^2]$

[0058] There is observed a significant performance gain provided by the proposed invention, especially in low SNR regime. It can be seen that the invention is indeed more able to cope with the noise impairment which is more thoroughly modelled than in Moose's algorithm. Note also that a wrongly assigned prior $p(\theta|l)$ on the channel realization does not

lead to critical performance decay; in the high SNR region, it is almost unimportant.

[0059] Figures 4-6 show the performance of the steepest descent algorithm which was described above. The system parameters are the same as in the previous simulation, with a correct prior $L_{\text{assumed}} = 3$ on the channel length at the receiver. The termination constraint is simply the number of iterations k of the inner loop, which we limit to $k=3$, $k=5$, $k=10$ and $k=50$. It is observed that saturations appear for $k < +\infty$, which are explained by the systematic error introduced by the minimal step size 2^{-k} in the iteration loop. For $k > 10$, the performance plots (which we did not provide for clarity) fit the plot $k=50$ in the -15 dB to 10 dB SNR range. Note also that the saturated standard deviation (defined as $E[(\hat{\theta}-\theta)^2]^{1/2}$) for $k=5$ is around 1% of the subcarrier spacing, which corresponds to the maximum allowable CFO mismatch in most OFDM systems. Therefore, 5 iterations might be sufficient to ensure a reliable estimation of the CFO.

[0060] Figure 7 illustrates the impact of the choice of the particular pilot sequence for executing the process of the invention. Moose's randomly generated double-half pilot sequences as well as QPSK random sequences are compared against the primary and secondary synchronization sequences (PSS, SSS) from the 3GPP Long Term Evolution standard. There is observed a large performance difference between those two types of pilots. This is simply due to the fact that both PSS and SSS are not of constant modulus over time; this makes part of the signal more sensible to noise and part of the signal less sensible to noise, but in average, this leads to less efficient pilots in terms of CFO estimation. It should be noticed that also Moose's sequence is in no way better than any randomly generated sequence, which demystifies the original insightful idea from Moose.

Applications of the invention

[0061] This invention fits typically the needs of the 3GPP-LTE standard for which no sequence dedicated to CFO estimation is provided. Due to its generality, this method can be applied in many OFDM systems which seek for CFO estimation while not having access to the channel information. Since this scheme has a complexity which scales with the number of iterations of the algorithm, it can be adapted to rough low consumption estimates at the receiver as well as thin higher consumption estimates at the base station.

[0062] The invention is particularly adapted to the Long Term Evolution standard, during the PSS to SSS synchronization phases.

[0063] The invention provides CFO estimation process prior to channel estimation in OFDM for any available pilot sequence. This is a very advantageous effect which was not known with prior art techniques: usual CFO techniques come along with a dedicated sequence. With the new technique which is proposed, there is no need of any specific sequence. Furthermore, it has been observed that the process is particularly effective when the sequence is composed of symbols having constant amplitude. It is then more advantageous to run this method on the most appropriate pilots.

[0064] This invention eliminates the problem of initial synchronization prior to channel estimation. It can also help estimating the CFO from signals coming from interfering base stations whose channels have not been estimated. It is believed that no such general pilot-independent scheme has ever been proposed in the OFDM contest.

[0065] Furthermore, it has been observed that the technique described above shows better performance than the classical ad-hoc techniques based on the first derivations of Moose. In the maximum a posteriori performance viewpoint, it has even been proved that that technique is optimal.

Claims

1. Process for estimating the carrier frequency offset (CFO) for a Orthogonal Frequency Division Multiplex (OFDM) communication system, said process being performed in a receiver receiving a pilot sequence $\mathbf{x} = (x_0, x_1, x_2, \dots, x_{N-1})^T$, said process involving the step of determining a value of θ which maximizes the formula:

$$C(\theta) = -\mathbf{d}_\theta^H \mathbf{Y}^H \mathbf{X} (\mathbf{X}^H \mathbf{X} + \sigma^2 \mathbf{Q}^{-1})^{-1} \frac{1}{\sigma^2} \mathbf{X}^H \mathbf{Y} \mathbf{d}_\theta$$

where

$$\mathbf{X} = \begin{pmatrix} x_0 & x_{N-1} & \cdots & x_{N-L-1} \\ x_1 & x_0 & \cdots & x_{N-L-2} \\ \vdots & \vdots & \vdots & \vdots \\ x_{L-2} & x_{L-3} & \cdots & x_{N-1} \\ x_{L-1} & x_{L-2} & \cdots & x_0 \\ \vdots & \vdots & \vdots & \vdots \\ x_{N-1} & x_{N-2} & \cdots & x_{N-L} \end{pmatrix}$$

$$\mathbf{d}_\theta = (1, e^{2\pi \frac{i\theta}{N}}, \dots, e^{2\pi \frac{(N-1)i\theta}{N}})^T;$$

$\mathbf{Y}$ is the diagonal matrix of main diagonal $\mathbf{y}$;
 σ^2 is the noise power and $\mathbf{Q}$ is the channel time covariance matrix.

2. Process according to claim 1 **characterized in that** said matrix $\mathbf{Q}$ is set to be equal to $1/L \mathbf{I}_L$, with $\mathbf{I}_L$ being the $L \times L$ identity matrix and L corresponding to the presumed length of the channel.
3. Process according to claim 1 or 2 **characterized in that** a dichotomy process is used for estimating the value of θ Which maximizes $C(\theta)$.
4. Process according to claim 1 **characterized in that** said matrix $\mathbf{Q}$ is contextually determined from the physical location of the receiver.
5. Process according to anyone of the preceding claims **characterized in that** it involves the steps of :
 - receiving (21) an input signal $\mathbf{y} = (y_0, \dots, y_{N-1})^T$;
 - detecting (22) said pilot signal;
 - computing (23) an estimation of the variance of the noise σ^2 ;
 - setting an estimation (24) of the channel time covariance matrix $\mathbf{Q}$;
 - computing elements $A_{n,m}$ of a $\mathbf{A}$ matrix defined as follows:

$$\mathbf{A} = \mathbf{Y}^H \mathbf{X} (\mathbf{X}^H \mathbf{X} + \sigma^2 \mathbf{Q}^{-1})^{-1} \frac{1}{\sigma^2} \mathbf{X}^H \mathbf{Y}$$

where $\mathbf{X}$ is a pseudo - circulant matrix defined from said pilot signal
 $\mathbf{x} = (x_0, x_1, x_2, \dots, x_{N-1})^T$ as follows :

$$\mathbf{X} = \begin{pmatrix} x_0 & x_{N-1} & \cdots & x_{N-L-1} \\ x_1 & x_0 & \cdots & x_{N-L-2} \\ \vdots & \vdots & \vdots & \vdots \\ x_{L-2} & x_{L-3} & \cdots & x_{N-1} \\ x_{L-1} & x_{L-2} & \cdots & x_0 \\ \vdots & \vdots & \vdots & \vdots \\ x_{N-1} & x_{N-2} & \cdots & x_{N-L} \end{pmatrix}$$

$\mathbf{Y}$ is the diagonal matrix of main diagonal $\mathbf{y}$;

σ^2 is the noise power .

- computing (26) the $N-1$ values of $\hat{a}_k$ in accordance with the formula:

$$\tilde{a}_k = \sum_{m=1}^{N-k} A_{k+m,m}$$

- computing (27), the following vectors:

$$\tilde{\mathbf{b}}_{\Re}^T = \Re[\tilde{a}_1, 2\tilde{a}_2, \dots, (N-1)\tilde{a}_{N-1}]$$

$$\tilde{\mathbf{b}}_{\Im}^T = \Im[\tilde{a}_1, 2\tilde{a}_2, \dots, (N-1)\tilde{a}_{N-1}]$$

- initializing (29) a loop with the two variables:

$$\theta_{\min} = -1/2$$

and

$$\theta_{\max} = 1/2$$

- computing (30) the value of

$$d = \tilde{\mathbf{b}}_{\Re}^T \cdot \mathbf{s}_{\theta} + \tilde{\mathbf{b}}_{\Im}^T \cdot \mathbf{c}_{\theta}$$

with

$$\mathbf{c}_{\theta}^T = [\cos(2\pi\theta/N), \dots, \cos(2\pi(N-1)\theta/N)]$$

$$\mathbf{s}_{\theta}^T = [\sin(2\pi\theta/N), \dots, \sin(2\pi(N-1)\theta/N)]$$

- testing (31) the sign of d ;

- if d is positive, then updating (32) the value of $\theta_{\min}$ as follows:

$$\theta_{\min} = (\theta_{\min} + \theta_{\max})/2$$

- if d is negative, then updating (33) the value of $\theta_{\max}$ as follows:

$$\theta_{\max} = (\theta_{\min} + \theta_{\max})/2$$

- repeating step 29-33 until the completion of said loop being determined by any predefined stopping requirement);

- completing (35) said process and returning an estimation of said CFO in accordance with the formula:

$$\hat{\theta} = [\theta_{\min} + \theta_{\max}]/2$$

6. Process according to claim 5 **characterized in that** only the upper right coefficients $A_{n,m}$, with $n>m$, are being computed during step 26.

7. Process according to anyone of the preceding claims **characterized in that** said pilot sequence is a Primary Synchronization Signal.

8. Process according to anyone of the preceding claims **characterized in that** said receiver is a 3GPP-LTE communication system.

9. Apparatus for an Orthogonal Frequency Division Multiplexing communication system comprising:

- a receiver for receiving an input signal comprising a pilot sequence $\mathbf{x} = (x_0, x_1, x_2, \dots, x_{N-1})^T$,
- means for determining a value of θ which maximizes the formula:

$$C(\theta) = -\mathbf{d}_{\theta}^H \mathbf{Y}^H \mathbf{X} (\mathbf{X}^H \mathbf{X} + \sigma^2 \mathbf{Q}^{-1})^{-1} \frac{1}{\sigma^2} \mathbf{X}^H \mathbf{Y} \mathbf{d}_{\theta}$$

where

$$\mathbf{X} = \begin{pmatrix} x_0 & x_{N-1} & \cdots & x_{N-L-1} \\ x_1 & x_0 & \cdots & x_{N-L-2} \\ \vdots & \vdots & \ddots & \vdots \\ x_{L-2} & x_{L-3} & \cdots & x_{N-1} \\ x_{L-1} & x_{L-2} & \cdots & x_0 \\ \vdots & \vdots & \ddots & \vdots \\ x_{N-1} & x_{N-2} & \cdots & x_{N-L} \end{pmatrix}$$

$$\mathbf{d}_{\theta} = (1, e^{2\pi \frac{i\theta}{N}}, \dots, e^{2\pi \frac{(N-1)i\theta}{N}})^T;$$

$\mathbf{Y}$ is the diagonal matrix of main diagonal $\mathbf{y}$;

σ^2 is the noise power and $\mathbf{Q}$ is the channel time covariance matrix.

10. Apparatus according to claim 9 **characterized in that** said matrix $\mathbf{Q}$ is set to be equal to $1/L \mathbf{I}_L$, with $\mathbf{I}_L$ being the $L \times L$ identity matrix and L corresponding to the presumed length of the channel.

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11. Apparatus according to claim 9 or 10 **characterized in that** it includes means for performing a dichotomic computation of the value of θ which maximizes $C(\theta)$.

5 12. Apparatus according to anyone of claims 9-12 **characterized in that** said pilot sequence is a Primary Synchronization Signal.

13. Apparatus according to anyone of the preceding claims **characterized in that** said receiver is a 3GPP-LTE communication system.

10 14. User Equipment comprising means for performing the process as defined by anyone of claims 1-8.

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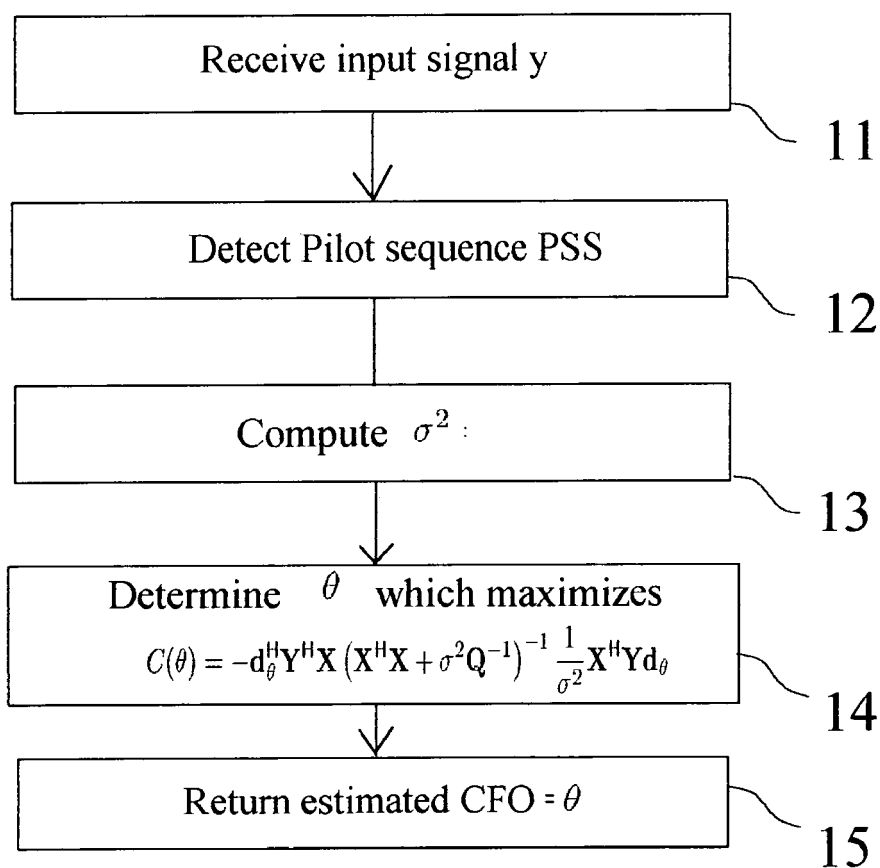


Fig. 1

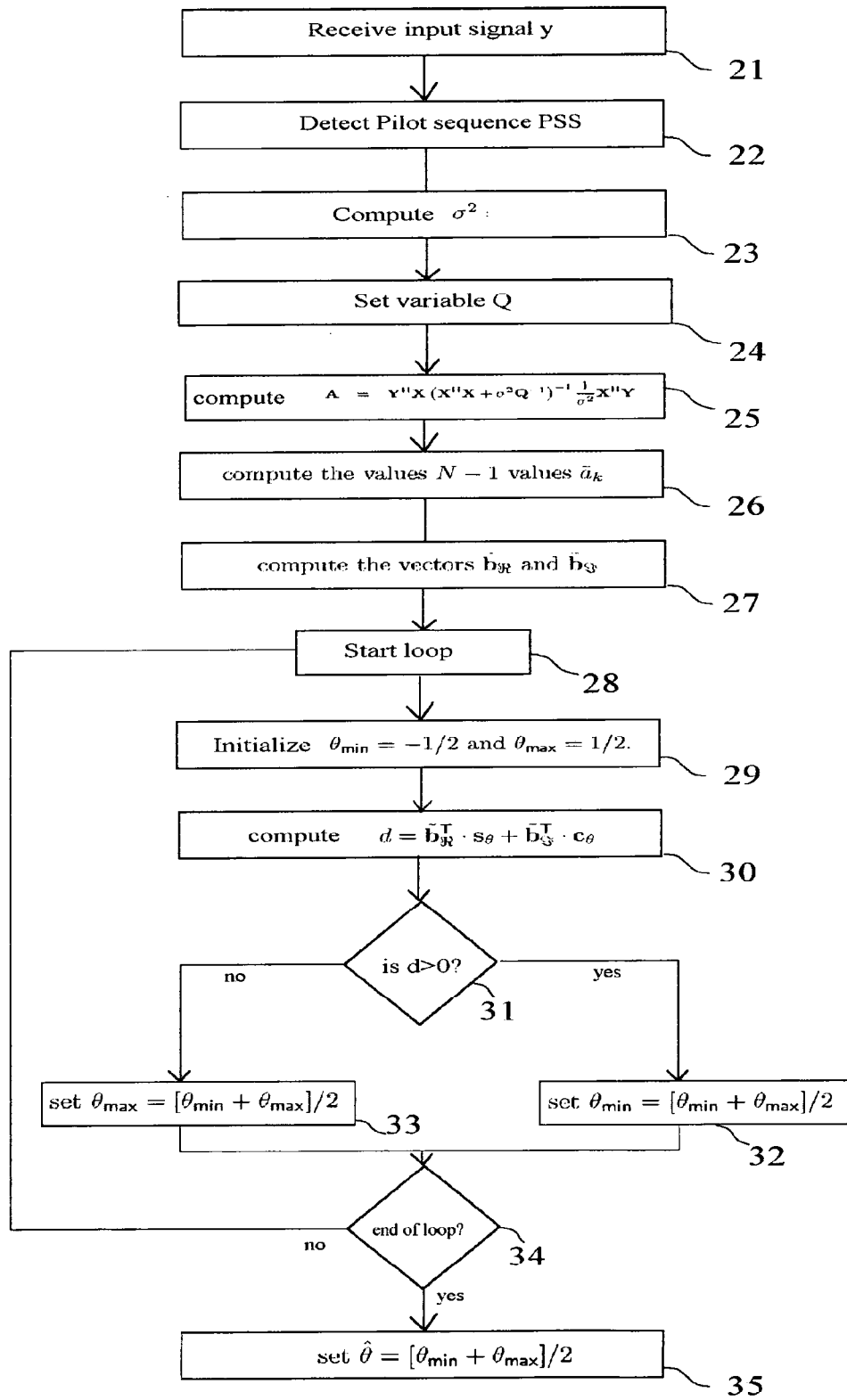


Fig. 2

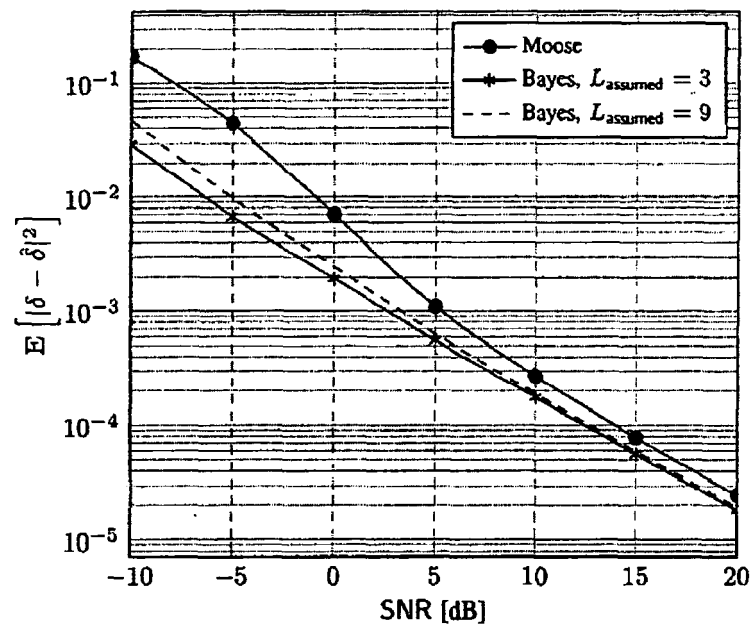


Fig. 3

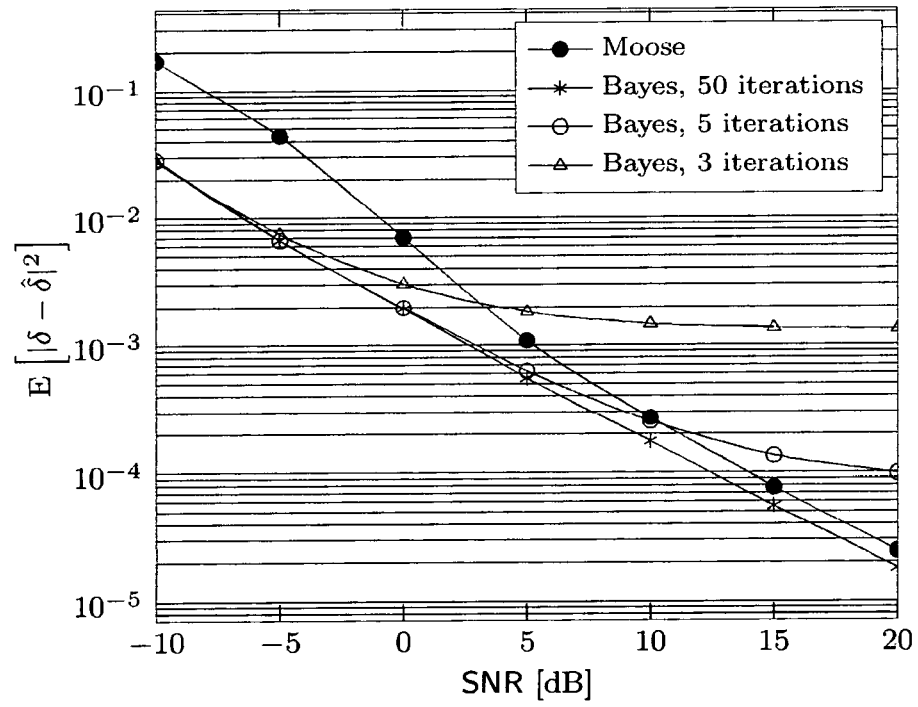


Fig. 4

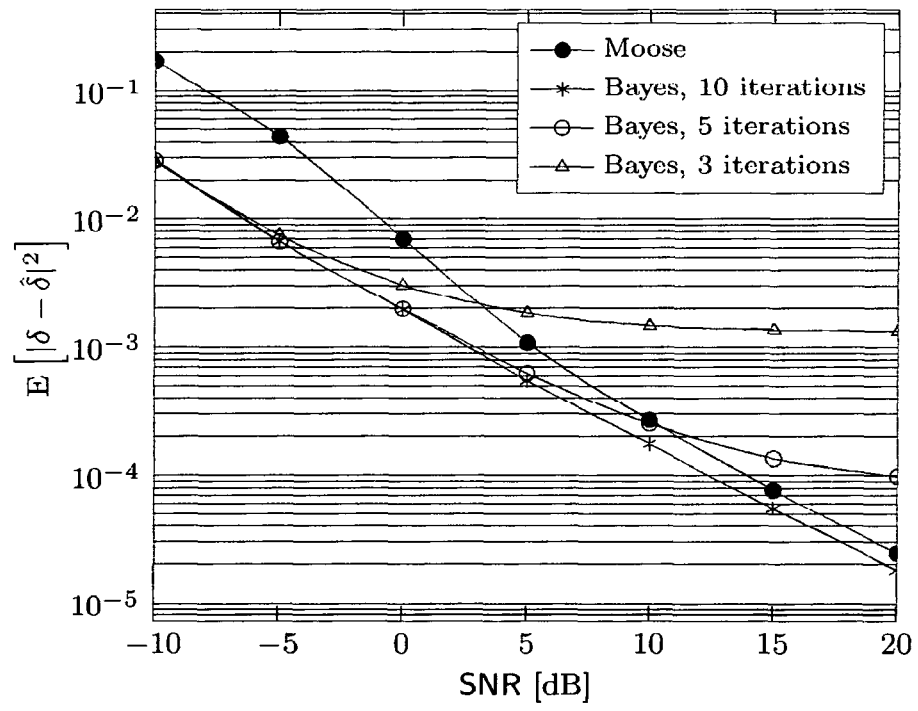


Fig. 5

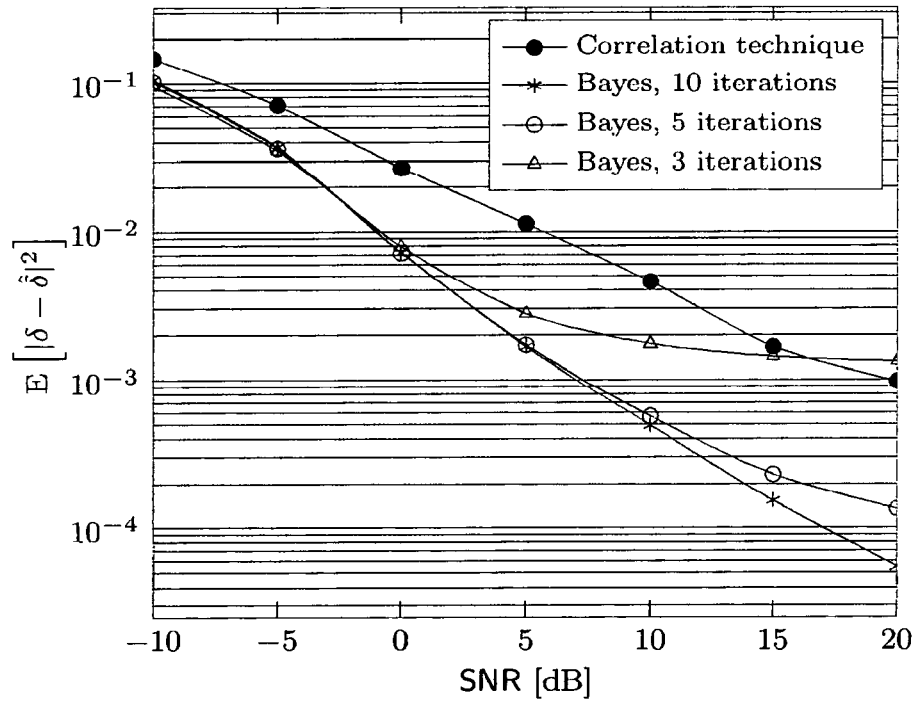


Fig. 6

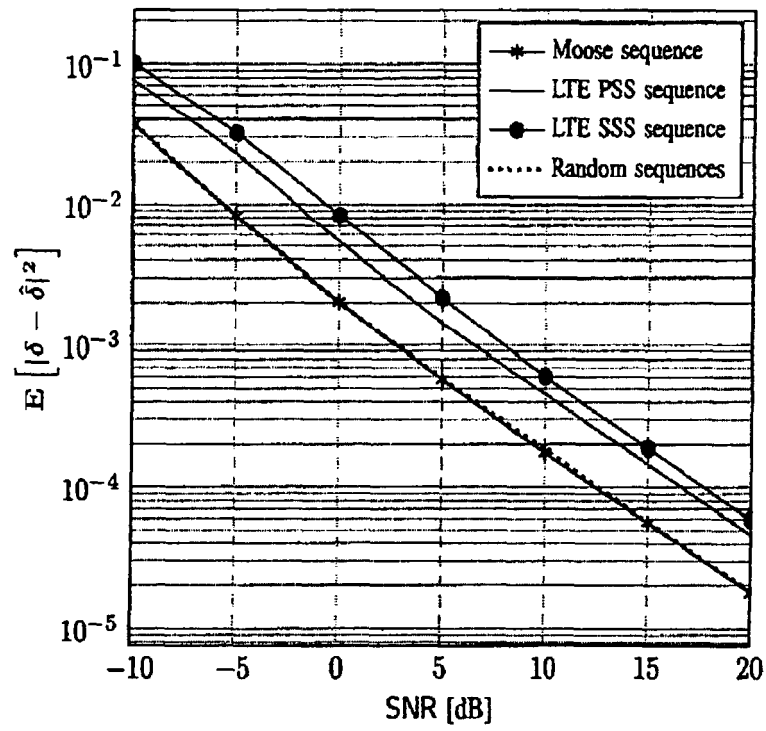


Fig. 7

REFERENCES CITED IN THE DESCRIPTION

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Non-patent literature cited in the description

- **P. H. Moose.** A technique for orthogonal frequency-division multiplexing frequency offset correction. *IEEE Trans. on Communications*, October 1994, vol. 42 (10), 2908-2914 **[0005]**